Amendments to the Specification:

Please replace the paragraph beginning at page 1, line 5 with the following amended paragraph:

-- This invention relates to [[a]] Digital Subscriber Line (DSL) transmission systems employing a ~~Discreet~~ Discrete Multitone (DMT) modulation scheme and, more particularly, to ~~a system~~ receivers and ~~method~~ methods for ~~optimizing bit allocation in such an implementation~~ use in such systems and schemes.

Please replace the paragraph beginning at page 1, line 18 with the following amended paragraph:

-- The ~~Discreet~~ Discrete Multitone (DMT) modulation technique is a particularly suitable way of implementing DSL in order to provide high speed and efficient use of existing bandwidth. In DMT, multiple narrow bandwidth ~~carriers~~ sub-carriers all transmit at once in parallel and each carry a small fraction of the total information. The sub-carriers or tones, each corresponding to a subchannel, have different center frequencies, ~~discreet bands or subchannels~~ are independently modulated with data ~~a carrier frequency corresponding to the center frequency of the carriers~~ and are processed in parallel. For example, in DMT, the bandwidth may be divided [[into]] between 256 sub-carriers, each with a 4 kHz bandwidth. Multi-carrier modulation requires orthogonality between all the sub-carriers and a Fast Fourier Transform (FFT) is a convenient method of achieving this modulation. At low frequencies, where copper wire attenuation is low and signal to noise ratio (SNR) is good, it is common to use a high bit rate, but at higher frequencies when unfavourable line conditions exist, modulation is relaxed to accommodate lower signal to noise ratio. Impulse noise which may be generated by electrical appliances, lightning or with a phone going off hook or ringing, is wideband in frequency and narrow in time so it is spread over many DMT subchannels and its influence over any subchannel is relatively small. Nonetheless, the line impulse response can result in symbol distortion at the receiver. In order to offset the line impulse response, it is known to insert a cyclic prefix (M) in the time domain samples. The cyclic prefix between symbols tends to reduce inter-symbol interference. Frequently, however, the line impulse response will last longer than the typical cyclic prefix time samples and, to address this, it is known to implement a time domain channel equalizer (TEQ) algorithm. The TEQ algorithm, which applies a Fast Fourier Transform, attempts to shorten the line impulse response into M time samples. However, the response can never be exactly the finite duration of the M ~~sample~~ samples regardless of the kind of TEQ algorithm implemented. As a result, there is always channel leakage outside the M samples which causes symbol distortion or inter-channel interference at the receiver. Although the interference level is mostly very small, the interference is not evenly distributed about the channels and some channels

may experience much higher interference which will cause receiver data error. ~~In the present invention, the cross-channel interference with imperfect or inadequate TEQ is estimated and the transmitted data allocation is based on the interference levels. For example, more bits are provided to the channel with less interference, and less bits are provided to the channel with high interference. This controls the bit error rate into the desired level.~~ --

Please insert the following paragraph at page 2, line 27:

-- In embodiments of the present invention, the cross-channel interference with imperfect or inadequate TEQ is estimated and the transmitted data allocation is based on the interference levels. For example, more bits are provided to the channel with less interference, and fewer bits are provided to the channel with high interference. This controls the bit error rate to the desired level. --

Please replace the paragraph beginning at page 2, line 28 with the following amended paragraph:

-- Therefore, in accordance with a first aspect of the present invention there is a provided a method of determining interference between channels in a digital subscriber line (DSL) transmission system employing ~~discreet~~ discrete multitone (DMT) modulation comprising the steps of: –

determining a peak power mask level per channel (Pk);

obtaining a channel impulse [[~~value~~]] response (h(n)) after implementation of a time equalization (TEQ) algorithm;

zeroing M main coefficient values of the channel impulse response (h(n)) to produce a residual impulse response (h′(n));

obtaining from the residual impulse response (h′(n)) a corresponding residual impulse spectrum (H′(k)); and

multiplying the per channel power mask level (Pk) and [[~~a~~]] residual impulse spectrum (H′(k)) to obtain a cross-channel interference level.

Please replace the paragraph beginning at page 3, line 7 with the following amended paragraph:

-- In accordance with a second aspect of the present invention there is provided a method of estimating cross-channel interference (I(k)) in a ~~discreet~~ discrete multitone (DMT) communication system implemented in a digital subscriber line (DSL) application, the DMT system employing inter-symbol cyclic prefix (M) and time equalization (TEQ), the method comprising:

(a) measuring a total channel impulse response (h(n)) value remaining after TEQ;

(b) zeroing M main coefficients from h(n) to produce a residual impulse response (h′(n));

(c) performing fast Fourier transform (FFT) analysis on the ~~result of~~ residual impulse response

(h'(n)) ~~step (b)~~ to provide a residual impulse spectrum (H'(k)); and

(d) obtaining I(k) by multiplying the ~~result of~~ residual impulse spectrum (H'(k)) ~~step (c)~~ with a measured maximum power per channel value.

Please replace the paragraph beginning at page 4, line 4 with the following amended paragraph:

-- [[~~The~~]] Preferred embodiments of the invention will now be described in greater detail and by way of example only, with reference to the attached drawings wherein:

Fig. 1 shows the frequency and time relationship in a ~~discreet~~ discrete multitone (DMT) implementation;

Fig. 2 is a block diagram of a basic DMT transmitter;

Fig. 3 is a basic block diagram of a DMT receiver;

Fig. 4 shows at a high level the bit allocation process; and

Fig. 5 illustrates at a high level the interference measurement process. --

Please replace the sub-heading and paragraph beginning at page 4, line 13 with the following amended sub-heading and paragraph:

-- Detailed Description of Preferred Embodiments of the Invention

As discussed previously, DMT operates in the frequency domain. Each DMT subchannel (there are many being communicated in [[~~the~~]] parallel) lasts a long time but operates in a narrow frequency band of the total bandwidth. As shown in FIG. 1, the total ~~carrier~~ frequency band is divided into many narrow bands, each corresponding to a sub-carrier or sub-channel, and each sub-carrier is modulated with information in the time domain. Bits are allocated to each subchannel for transmission and, at the receiver, the bits in each subchannel are demodulated to retrieve the communicated information. --

Please replace the paragraph beginning at page 4, line 22 with the following amended paragraph:

-- In the DMT based DSL system N point FFT is used to transform N frequency subchannel carriers having quadrature amplitude modulation (QAM) modulated thereon into N point time domain samples. Thus in the DMT transmitter shown in FIG. 2, input data is provided to a serial to parallel encoder. The output has N QAM symbols imparted thereto and, following the Inverse Fast Fourier Transform (IFFT) 12, the data is broken into time domain samples and, following the parallel to serial converter 14 the time domain signal may be passed through a digital to analogue converter (not shown) and on to the output line. Typically a cyclic prefix (M) [[~~16~~]] is inserted into the time domain

signal prior to transmission~~.~~, as indicated by box 16.

Please replace the paragraph beginning at page 5, line 1 with the following amended paragraph:

-- ~~The received signal as shown in~~ Referring now to FIG. 3, in the DMT receiver, the received signal is passed through an A to D converter (not shown). Typically, the cyclic prefix M has a smaller value than N for minimum redundancy. Because most line impulse responses [[~~lasts~~]] last much longer than M time samples, a time domain channel equalizer (TEQ) 18 is required. There are several TEQ algorithms available that are used to shorten the channel into M time samples. However, the channel can never have an exactly finite duration with only M samples, no matter what kind of TEQ algorithm is applied. As a result, there is always channel leakage outside the M samples which will cause symbol distortion or inter-channel interference at the receiver. Although the interference level is usually very small, the interference is not evenly distributed between the channels, and some channels may experience much higher interference levels which, in turn, will cause receiver data error. --

Please replace the paragraph beginning at page 5, line 13 with the following amended paragraph:

-- FIG. 4 shows ~~pictorially~~ schematically the bit ~~algorithm~~ allocation scheme, including cross-channel interference measurement according to the present invention. Typically, the bit allocation algorithms use the following information to decide how many bits should be transmitted on each channel: power mask or maximum transmission power at each channel; noise level including near end cross talk (NEXT) and far end cross talk (FEXT), the signal to noise ratio (SNR) requirements for bit transmission and the channel impulse response which gives channel attenuation for each channel. The output of the bit allocation algorithm provides: the number of bits to be transmitted per channel and the corresponding transmitting power for each channel. --

Please replace the paragraph beginning at page 5, line 23 with the following amended paragraph:

-- In FIG. 4, the first bit allocation algorithm 20 takes inputs [[~~N(k)~~]] V(k) which is the measured noise power for each channel; P(k) which is the power mask level which limits ~~maximum~~ transmission power for each channel; and SNR(bn) which is the signal to noise ratio requirement when bn bits are transmitted at the given channel. As shown in FIG. 4, the outputs of the first bit algorithm are b(k) which is the number of bits to be transmitted [[~~at~~]] via channel k in accordance with the bit allocation algorithm 20 and Px(k) which is the transmission power for channel k. --

Please replace the paragraph beginning at page 6, line 1 with the following amended paragraph:

-- The interference measurement block 22 of FIG. 4 measures the cross-channel interference level I(k) based on the total channel impulse response h(n) after the time equalization (TEQ) algorithm and the transmission power input [[~~P(k)~~]] Px(k) from the first allocation algorithm. The interference I(k) which is the statistically combined interference to channel k from all other channels ~~will be~~ is added to the measured noise level V(k) and the result $V_1(k)$ becomes the modified noise power input which is provided to the second bit allocation algorithm 24. The interference measurement algorithm, vis. block 22, also outputs the new power mask $P_1(k)$ for the second bit allocation algorithm 24. The reason for a new power mask function is that the cross-channel interference I(k) is signal dependent. Since the total noise level increases with cross-channel interference noise being added, the signal transmitting power will be increased as well. Consequently, the cross-channel interference I(k) will be increased which otherwise would result in a repeating loop. The interference measurement algorithm, block 22, calculates the new power mask $P_1(k)$ based on the pre-calculated transmitter power Px(k) and the maximum power transmitted will be fixed at $P_1(k)$. The calculated interference is based on the maximum possible transmission power $P_1(k)$. With the new power mask $P_1(k)$ and noise power $V_1(k)$, the second bit allocation algorithm 24 is implemented to obtain the final bit allocation [[~~b1(k)~~]] $b_1(k)$ and its corresponding transmitting power $Px_1(k)$ which takes into account the aforementioned cross-channel interference. --

Please replace the paragraph beginning at page 6, line 20 with the following amended paragraph:

-- One way to determine the new power mask function is described as follows:

Let ~~Pmax = max (Px(k))~~ P_{max} = max (Px(k)), then we have

$$P_1(k) = \begin{cases} P_{max}, & P(k) \geq P_{max} \\ P(k), & P(k) < P_{max} \end{cases}$$

Please replace the paragraph beginning at page 6, line 25 with the following amended paragraph:

-- The above equation means that [[~~Pmax~~]] P_{max} is chosen as the new maximum transmission power for all channels unless it is above the original mask power, in which case the original mask power P(k) is retained. In theory, with the newly calculated mask power, it is possible to calculate ~~statistic~~ statistical interference [[~~level~~]] levels in all channels. However, the precise calculation is still a very complicated procedure that involves the computation of each channel interference from all other channels. FIG. 5 shows a simple approximate upbound or upper limit for the cross-channel

interference with given total channel impulse response h(n) after TEQ.

Please replace the paragraph beginning at page 7, line 4 with the following amended paragraph:

-- As mentioned before, with an M point cyclic prefix being added in the ~~transmit~~ transmitted signal, there will be no interference if the total channel response h(n) is of finite length lasting only M sample points. The block "zeroing M coefficients" identified by reference numeral 26 in FIG. 5 sets those M coefficients to zeros and the remaining coefficients in the resulting residual impulse response h'(n) [[~~h(n)~~]] ~~are then the source of~~ then represent the inter-channel interference. FFT unit 28 takes the ~~Taking~~ Fourier transform ~~in FFT block 28~~ of the ~~remaining~~ residual channel impulse response h'(n) to obtain a residual impulse spectrum H'(k) and ~~multipling them at~~ multiplier 30 multiplies the resulting residual impulse spectrum H'(k) with $P_1(k)$ to give [[~~gives~~]] out the upbound for the cross-channel interference I(k).

Please replace the paragraph beginning at page 7, line 12 with the following amended paragraph:

-- To get the interference level, it is possible to use either I(k) or the ~~envelop~~ envelope of I(k), or some other ~~modificatied~~ modified form of I(k). [[~~$P_1(K)$~~]] $P_1(k)$ can also be modified based on Px(k) and P(k). In the $P_1(k)$ calculation, Pmax can be replaced with a local maximum of Px(k), such that ~~Pmax(k) = maxk1(Px(k+k1)), for $-C<k1<C$,~~ $P_{max}(k) = max_{k1}(Px(k+k1))$, for $-C<k1<C$ where C is predetermined value to decide the neighboring area around K. If $P_1(k)$ is chosen such that $P_1(k)$ = min([SNR(max(bn))][max(V(k))], max(P(k))), the first bit allocation in FIG. 4 becomes unnecessary. In this way, one half the computation for bit allocation is saved by running it only once and $P_1(k)$ becomes a constant which is the maximum potential transmitted power at each channel. --